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Transistor Amplifiers for Sound Broadcasting

by

S.D. BERRY, Associate I.E.E.
(Designs Department, BBC Engineering Division)

BRITISH BROADCASTING CORPORATION

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FOREWORD

THIS is one of a series of Engineering Monographs published by the British Broadcasting Corporation. About six are produced every year, each dealing with a technical subject within the field of television and sound broadcasting. Each Monograph describes work that has been done by the Engineering Division of the BBC and includes, where appropriate, a survey of earlier work on the same subject. From time to time the series may include selected reprints of articles by BBC authors that have appeared in technical journals. Papers dealing with general engineering developments in broadcasting may also be included occasionally.

This series should be of interest and value to engineers engaged in the fields of broadcasting and of telecommunications generally.

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TRANSISTOR AMPLIFIERS FOR SOUND BROADCASTING

SUMMARY

This monograph is concerned with the application of transistors to the audio-frequency amplifiers employed in sound broadcasting, for which a high standard of performance is required. The principles followed in the design of such amplifiers are discussed and five examples of various types are described, with performance details. The operating conditions of d.c. feedback pairs and of 'super-alpha' pairs are analysed and some numerical evaluations are given. Finally some conclusions are drawn regarding the use of transistors in high performance amplifiers of this nature.

1. Introduction

Much of the published work regarding the use of transistors at audio frequencies has, up to the present, been concerned with their use in apparatus of comparatively low technical standards. For such apparatus the small size and large power efficiency of transistors have a great appeal and these characteristics have generally been utilized to the full at the expense of electrical performance. Little attention appears to have been directed towards the use of transistors for high-quality sound amplifiers.

For sound broadcasting a high standard of performance is essential and this may readily be obtained from valve amplifiers. Nevertheless, transistors have a number of attractive features which may very favourably influence studio and control room equipment. Work has therefore been carried out by the BBC's Engineering Division to determine whether the known advantages of transistors may be combined with an adequately high electrical performance and to design a number of amplifiers with the characteristics necessary for various applications in the programme chain. This monograph describes some of these amplifiers and presents some general conclusions.

2. The Requirements of Broadcasting and Amplifier Types

Amplifiers for the sound broadcasting chain may be divided into two broad groups:

(*a*) terminal amplifiers, which comprise those used only at the source or ultimate destination of a programme, e.g. studio or outside broadcast (O.B.) microphone amplifiers; and

(*b*) chain amplifiers, which comprise those which may be used at many points in a long land-line chain, e.g. line-receiving and sending amplifiers. Since a number are used in tandem the performance requirements of amplifiers of the class (*b*) group are considerably higher than those of the (*a*) group, which are used singly.

Transistors offer particular advantages for portable, high-gain amplifiers of low power consumption such as are required for use on outside broadcasts. An amplifier of this type, together with a programme meter, has been designed and several prototypes have been on service trial since May 1957 with encouraging results. This amplifier and programme meter, together with others of both Group (*a*) and Group (*b*), are listed and described below.

1. Outside broadcast amplifier. This has been produced in two forms. In one, it is combined with the programme meter of 2, below, and other facilities of a minor nature to be a close copy externally of the existing OBA/9 portable O.B. amplifier[1] in order to make it compatible with current operational techniques. In the other form the amplifier alone, coded AM9/1, is constructed in a manner suitable for mounting in groups as needed in a mobile control room. The circuit and performance are the same in both forms, and only the AM9/1 will be considered further.
2. A programme meter conforming to the normal BBC standards of programme volume measurement.[2] It is coded ME12/1.
3. A microphone pre-amplifier of small size for use in studio control desks.
4. A line-sending amplifier. This has a low, fixed gain with a higher output power than the other amplifiers and is intended for sending programme to inter-regional land-lines. Its performance closely approaches that of the standard C/9 valve amplifier described elsewhere,[2] and it is mounted in a chassis of the same construction.
5. A line-receiving amplifier. This has an adjustable gain and is primarily intended for amplifying the equalized programme received from inter-regional land-lines. It closely resembles the standard GPA/4A valve amplifier[2] in performance and construction.

3. General Design Considerations

The transistors used in this series of amplifiers are readily available germanium p-n-p junction types manufactured in this country. The primary aim has been to produce high-quality amplifiers for the particular needs of sound broadcasting; high power-efficiency and small size of apparatus have been secondary considerations, although the nature of transistors has resulted in these features being obtained to some degree.

3.1. *Basic Circuit and Feedback Arrangements*

A single transistor driven hard produces a good deal of even order distortion, and the use of a push-pull pair for the output stage is almost essential if good linearity is to be obtained. Also, the output transformer may be made smaller and its design simplified because of the reduction of magnetic polarization obtained with the push-pull arrangement. An emitter coupled pair form a convenient and

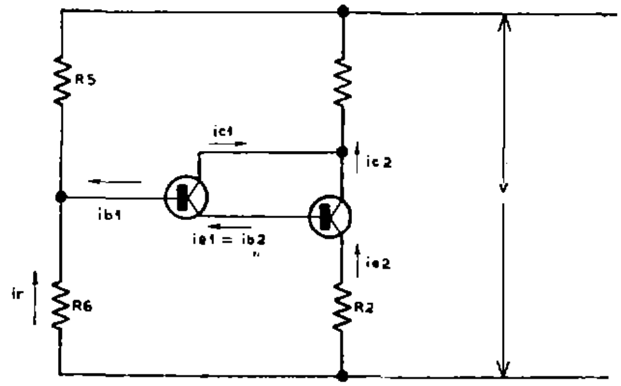

Fig. 1 — 'Super-Alpha' Pair

economical driver stage coupling the output transistors to the preceding single-ended stages; such a pair also produce a satisfactory gain and together with the output pair enable suitable a.c. and d.c. feedback loops to be contrived. Such twin pairs are used in the high-gain O.B. amplifiers (1) and also in the programme meter amplifier (2). For the microphone amplifier (3), where the output transistor is not driven hard and an output transformer is unnecessary, a single-ended output stage is adequate.

A fairly large amount of both a.c. and d.c. feedback must be used in amplifiers of this nature in order to obtain the good frequency response and low distortion required, to control the input and output impedances and to stabilize the d.c. operating points. In order to avoid a too great sacrifice of forward gain feedback should be taken over as many stages as possible but, with the variation found between individual specimens of the transistors at present available, the construction of a stable feedback loop of adequate gain over more than two stages is hardly practicable, unless each amplifier is individually adjusted or the transistors are selected to close tolerances. With feedback pairs, however, a reliable stability margin, large enough to accommodate all, except, perhaps, the most extreme specimens of transistors, may be readily obtained with some 20 dB of feedback. All the amplifiers are therefore built up, wherever possible, from feedback pairs.

For the output stages of the two amplifiers of Group 2, the line-sending and the line-receiving amplifiers, the arrangement of Fig. 1 is used in push-pull pairs. This interconnection of two transistors is attributed to Darlington and is described elsewhere.[3, 4] It behaves as a single transistor having a value of α given by $\alpha = 1-(1-\alpha_a)(1-\alpha_b)$ and the common emitter connection has therefore a current gain of $\alpha'_a\,\alpha'_b$ approximately, where a and b refer to each of the transistors of the pair. This high value of current gain enables a high input impedance to be obtained and an adequate amount of feedback to be used. Also, due to the compensatory effect of the first transistor output on that of the second, there is an improvement in linearity at high values of total output compared with the performance of the second transistor alone. These features are of particular value for the line-sending amplifier.

3.2. *Terminal Impedances and Load Conditions*

Constant input and output impedances of a specified magnitude are required for all amplifiers, and each must generally be controlled by feedback since the ratios of the input and output transformers are rigidly determined by considerations of noise factor and load impedance respectively.

With regard to the output circuit the optimum load impedance at each collector closely approaches the ratio of the direct voltage to the current of the transistor. Much less latitude is allowable in this respect with transistors than with valves if the maximum power output is to be obtained and the output transformer ratio is thus narrowly deter-

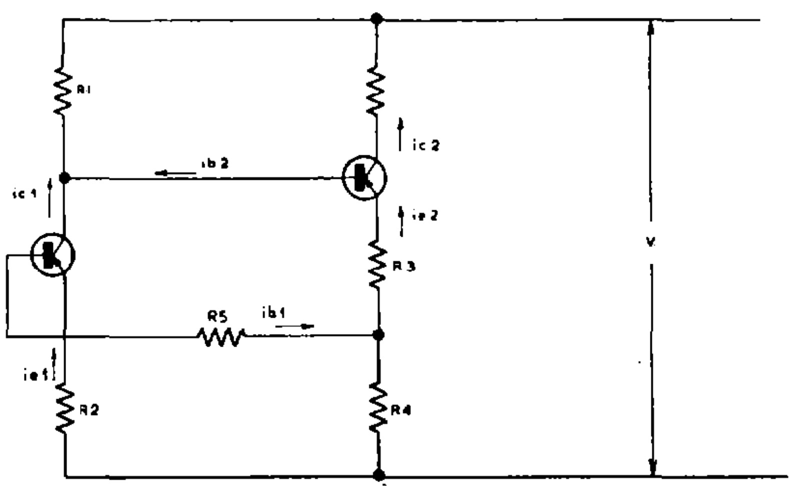

Fig. 2 — D.C. Feedback Pair

mined. Voltage negative feedback is then applied to drive the collector impedance down to a low value, which is padded out by series resistance to produce the required impedance at the amplifier output terminals.

With the transistor types available and a practicable degree of feedback the conflicting requirements of output impedance and optimum load can be met only if the dissipation of the output transistors is obtained by a high voltage and low current, rather than a low voltage and high current. For this reason, and also in order to obtain sufficient voltage for d.c. stabilization, a supply pressure of 24 V is used and the output transistors are run at the highest safe collector/emitter voltage though well within their permissible power dissipation.

The negative feedback from the output stage is obtained from a tertiary winding on the output transformer. The characteristics of this transformer have a good deal of influence on the performance and stability of push-pull feedback amplifiers of this type and considerable care is required in its design, particularly with regard to the symmetry and sectionalization of the windings, if the optimum performance is to be obtained from the amplifier. The output transformers of the amplifiers described have a core consisting of a $\frac{3}{4}$-in. butt-jointed stack of Telcon mumetal stampings, No. 39T.

3.3 *D.C. Bias Arrangements*

Since most of the transistor parameters are dependent upon operating conditions it is imperative that these latter are very well stabilized if a stable and reproduceable performance is to be obtained. This matter is particularly important for the input stage transistor and for the output stage pair. The collector current and collector/emitter voltage of the former are low, in order to obtain minimum noise, and little variation can be tolerated; the collector currents of the two output transistors must be stable and remain equal in order to maintain maximum power output conditions.

Where high d.c. stability is required the well-known arrangement of emitter resistance and potential divider base supply, while quite adequate in some cases, may entail an excessively high emitter resistance or an impracticably low resistance base supply in others. In the amplifiers described here use is made of d.c. coupled pairs with d.c. negative feedback overall. This arrangement has also been investigated and described[5] by Murray. The basic circuit is shown in Fig. 2. Great stability may be obtained by this means without the disadvantages of the simple scheme, compared with which it also shows other advantages. For example, a few less components are required and there is some economy of supply current. Also, since the potential of the driving stage collector may be nearly the same as the base potential of the output transistor to which it is coupled, it may often be inconvenient to ensure that an electrolytic coupling condenser would always have an adequate polarizing potential of the correct polarity. This difficulty is avoided with direct coupling.

An analysis of the circuit is given in the appendix where expressions for the collector current of each transistor are derived. Each expression contains factors representing the magnitude of the contribution due to the cut-off currents of both transistors. The co-efficients of these factors form the stability factors $S=\frac{di_c}{di_{co}}$, following Shea,[6] and a numerical evaluation is given later for the microphone pre-amplifier. It will be seen that for the second transistor one term is negative and partial cancellation occurs.

For the case of similar stages, emitter coupled through R2, similar expressions may be derived showing the magnitude of the cut-off current of all four transistors contributing to the total collector current of each. These expressions are given in the appendix—equations (9) and (10)—and numerical evaluations are made for the AM9/1 amplifier.

3.4 *Noise*

For the amplifiers which require the lowest possible noise factor the input transformer has a ratio such as to present an impedance of about 600 ohms at the base of the first transistor. The required input impedance to the amplifier is then obtained by a combination of both series and shunt connected negative feedback. It may be shown[7] that such a combination of opposite tendencies with regard to impedance is more effective in stabilizing the input impedance than either form acting alone. Moreover, by adjusting the relative proportions of each, a desired impedance may be obtained together with the total amount of feedback required by other considerations.

With the optimum impedance at the input and low collector current and voltage, a noise factor of 5 dB measured with a bandwidth of 400 c/s to 10 kc/s can be obtained with selected transistors of a low-noise type for the input stage. Without selection the noise factor may be about 7 dB. The contribution to the total noise of the second and even subsequent transistors is much more significant than would be the case in the corresponding valve amplifier due to the fact that it is usually impracticable to have the optimum noise conditions (e.g. low impedance in the base circuit and low collector current) for transistors after the first, and also because it may not be possible to use a low-noise transistor type. It has been found necessary, therefore, to pay attention to these matters and approach good noise conditions in the later stages as far as other considerations will permit in order to avoid a significant increase of total noise.

4. Circuit Details and Performance

4.1 *Outside Broadcasting Amplifier AM9/1*

This is a microphone-to-line amplifier of nearly 90-dB voltage gain primarily intended for use where a number of separate microphone circuits are required, each feeding to an individual line. This need arises, for example, when many commentators of differing nationalities are contributing in their various languages to an outside broadcast of world-wide interest. A slightly modified form, having a pre-set gain, has also been produced for use, in association with a mixing and control desk, in a mobile control-room.

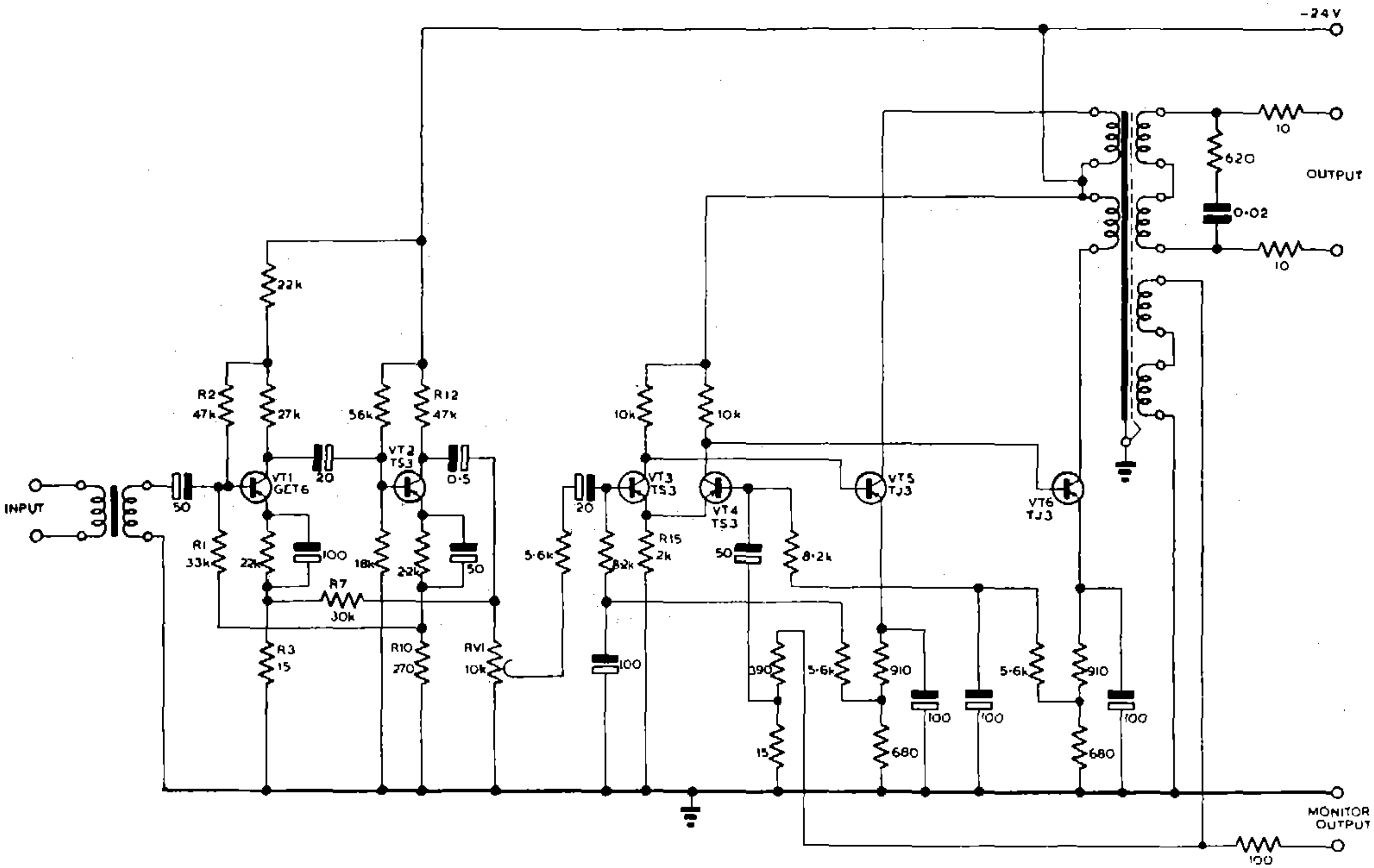

Fig. 3 — O.B. Amplifier: Circuit

The circuit of the AM9/1 is shown in Fig. 3. It will be seen that it is made up of two feedback pairs coupled by the gain control, RV1, a type of carbon-track logarithmic volume control which has good noise-free properties. It works between impedances of about 5 000 ohms and, in this condition, has suitable control and fading characteristics. The input amplifier has both shunt and series connected negative feedback applied to VT1 from R10/R1 and R7/R3 respectively. There is also some shunt-connected feedback over VT1 alone via the resistor R2 which, in conjunction with R1, also serves to supply bias current to VT1. The ratio of the input transformer is such as to convert the microphone impedance to 600 ohms at the base circuit of VT1 in order to obtain optimum noise conditions, and the two forms of feedback are adjusted in magnitude to give the required amplifier input impedance of 300 ohms. The total amount of feedback (defined here and subsequently as the reduction of the amplifier insertion gain by the feedback) is about 16 dB.

The d.c. bias arrangements of these transistors are conventional—the d.c. coupling given by R10 is small. The stability factors, S, are 1·5 for VT1 and 6·4 for VT2.

The output stage is a push-pull pair driven by the two transistors VT3 and VT4, which are coupled in anti-phase by means of the common emitter resistance R15. The interstage coupling is direct and the bias currents for each of the first transistors is supplied from a tapping on the emitter resistance of its output stage partner. There is therefore considerable d.c. feedback over each pair in cascade and each pair is d.c. coupled by R15. Expressions for the collector currents of the transistors resulting from this arrangement are given in the appendix. For this amplifier they become, in milliamps:

$$i_{c1}=1\cdot 13+2\cdot 9\, i_{co1}+0\cdot 7\, i_{co2}-0\cdot 6\, i_{co1a}-0\cdot 23\, i_{co2a}$$

$$i_{c2}=7\cdot 07-15\cdot 5\, i_{co1}+2\cdot 6\, i_{co2}+3\cdot 4\, i_{co1a}+1\cdot 3\, i_{co2a}$$

where i_c and i_{co} are collector current and cut-off current respectively and the suffixes 1, 2, 1*a*, and 2*a* denote respectively VT3 (or VT4), VT5 (or VT6), VT4 (or VT3), and VT6 (or VT5).

It will be seen that the collector currents are well stabilized; that of each output transistor is likely to fall slightly, and indeed does so, as the ambient temperature is raised. The α' of the output stage transistors has only a small influence on the collector currents; if they are, for example, 30 and 60 respectively, the collector currents will differ by only 5 per cent from this cause. The associated resistors have a $\pm$ 2 per cent tolerance and therefore the two collector currents will be equal to within 7 per cent or 0·5 mA even with the above values of α' and extreme resistor differences. In most cases less than half of this current difference will appear. This matter is of importance with regard to the design of the output transformer in ensuring a small resultant polarizing current.

Negative voltage feedback giving a gain reduction of 18 dB is applied to the base of VT4 from a tertiary winding on the output transformer. This winding also provides a low-impedance source for aural and visual monitoring of the amplifier output at a point sufficiently close to the output stage e.m.f. to be relatively unaffected by the impedance variations with respect to frequency of the line to which the output is connected.

Each sub-amplifier, excluding the gain control and output transformer, is mounted upon one of two printed-circuit cards and can therefore be easily replaced in the event of a fault.

Performance:

Source impedance 300 ohms.
Load impedance 240 ohms.
Maximum insertion gain at 1 kc/s: 87$\pm$1 dB.
Frequency response: 90 c/s to 10 kc/s between 0 and −0·5 dB,
60 c/s to 15 kc/s between 0 and −1·0 dB, relative to 1 kc/s

Non-linearity: Total harmonic distortion:
Power output +4 dBm: at 1,000 c/s: 0·15 per cent
at 60 c/s: 0·35 per cent
Power output +12 dBm (normal peak output):
at 1,000 c/s: 0·3 per cent
at 60 c/s: 0·7 per cent

Noise factor:
With a bandwidth of 400 c/s to 10 kc/s a noise factor of between 5 and 7 dB, depending upon the noise properties of the first transistor, may be readily obtained.

Terminal impedances at 1 kc/s:
$|Z \text{ out}|=75$ ohms$\pm$10 per cent
$|Z \text{ in}|=300$ ohms$\pm$10 per cent

Power consumption:
20 mA at 24 V.

4.2 *Programme Meter Amplifier ME*12/1

The meter used for the measurement of programme volume in BBC practice has a right-hand zero when passing no current. It is illustrated in Fig. 8. The pointer is deflected to the usual left-hand zero position when the associated amplifier-rectifier is switched on and the circuit is made to give a reduction of meter current with increasing signal amplitude. This arrangement is adopted in order to give the very quick meter response required. The meter is, in the standard valve-operated apparatus,[2] driven by a variable-mu valve, adjusted to give an approximately logarithmic scale shape over a range of 26 dB, preceded by a full-wave rectifier with a capacitative load having a charge time constant of 2·5 milliseconds and a discharge time constant of 1 second. These performance characteristics are also necessary in the transistor apparatus and the general arrangement of the valve apparatus is followed. That is, the signal is amplified and rectified into a resistance/capacitance load with the above time constants (C8 with R16 plus the input resistance of the following transistor) and applied to an approximately logarithmic d.c. stage in which is connected the normal type of indicating meter. The circuit is shown in Fig. 4.

In the design considerations for the final d.c. stage two difficulties exist which are not present where valves are used: a transistor is far more seriously affected by ambient temperature and the temperature effects of signal current

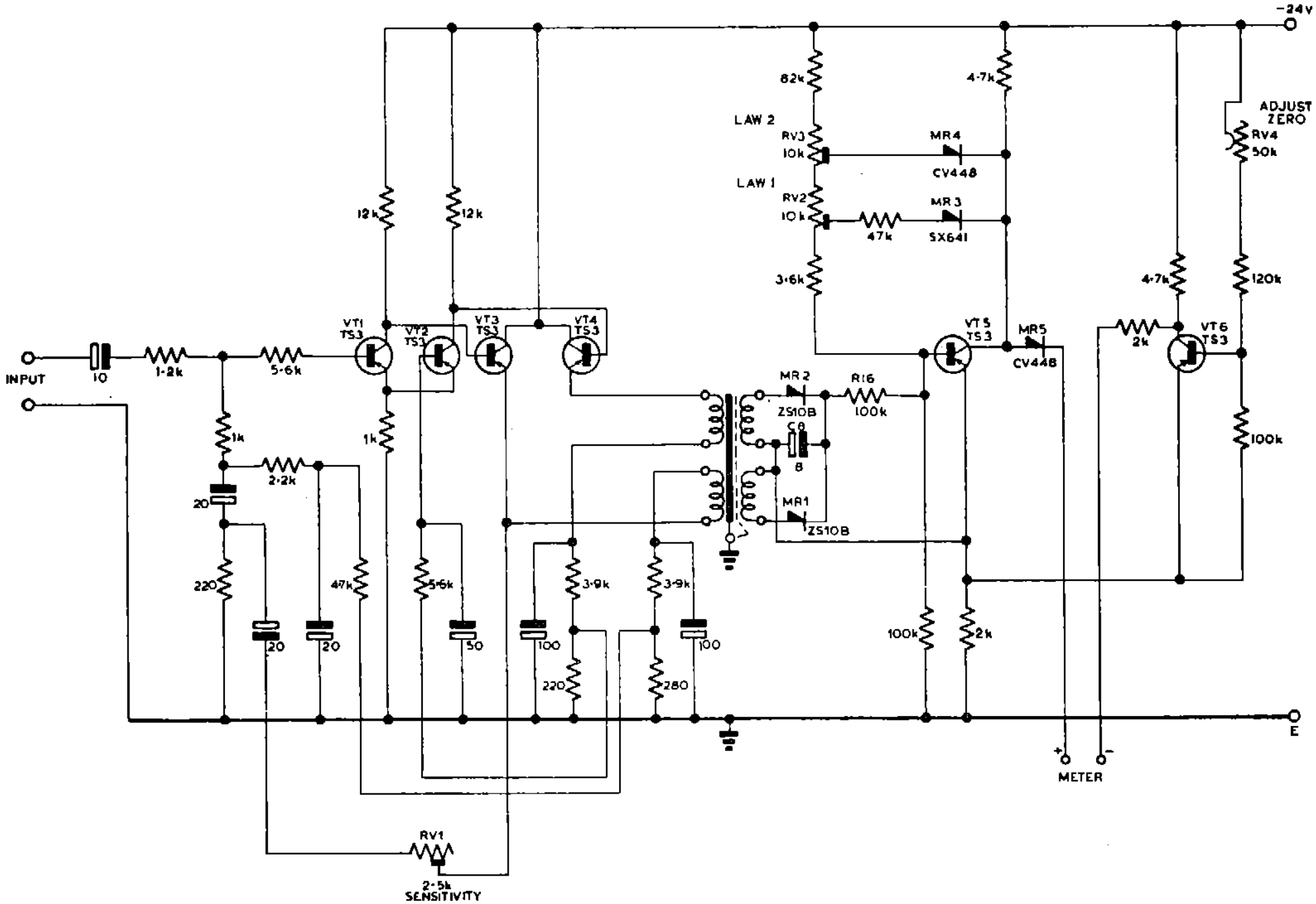

Fig. 4 — Programme Meter Amplifier: Circuit

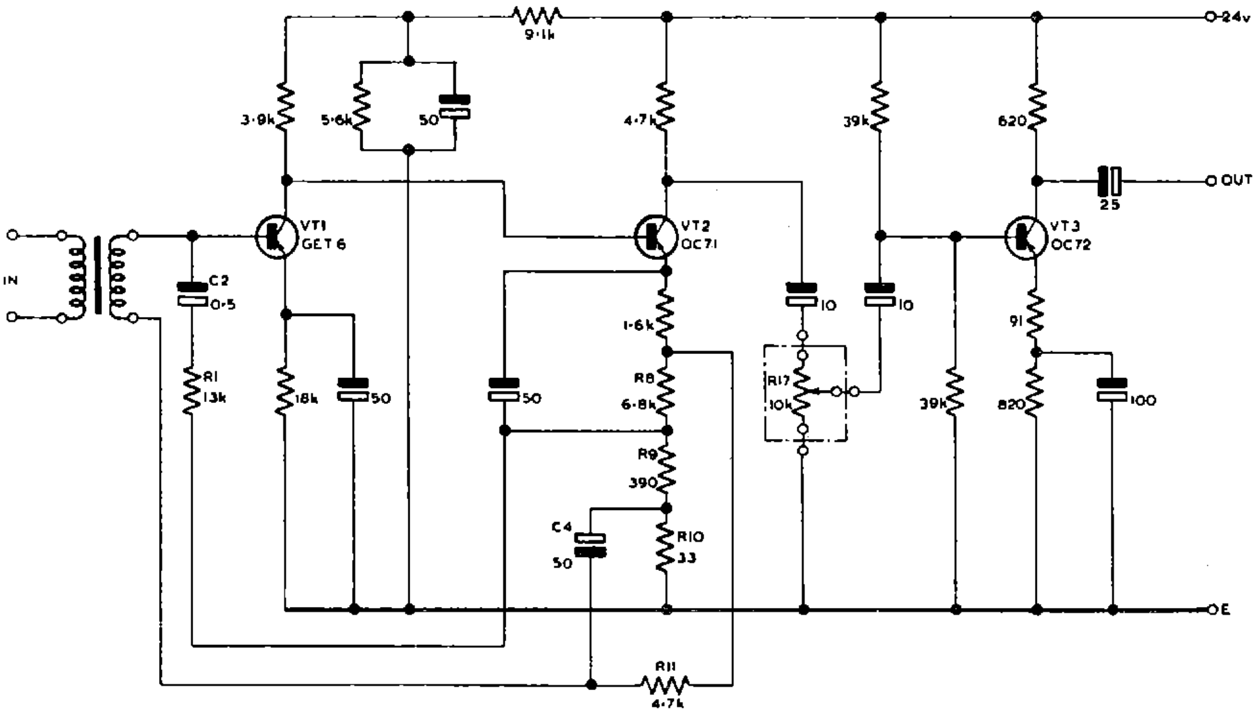

Fig. 5 — Microphone Pre-amplifier: Circuit

than is a valve, and also the long discharge time constant required cannot be obtained by the use of the high discharge resistance which a valve permits without a prohibitive reduction of sensitivity. Compared with the valve apparatus, therefore, a relatively large capacitance must be used in the time-constant circuit, and in order to obtain the short charging time a correspondingly low output impedance from the preceding amplifier is required. A satisfactory compromise is possible only when transistors of high-current gain (about 120) are used for the d.c. stage.

In order to minimize zero drift due to temperature changes the logarithmic d.c. stage is made up of the two transistors VT5 and VT6, emitter coupled, in the well-known bridge arrangement. This method gives satisfactory compensation for the effects of ambient temperature which acts on both transistors in a like manner. It does not necessarily, however, reduce drift caused by the junction temperature variations due to alterations of signal level, an effect which manifests itself as a slow creep of meter reading when a steady signal is applied. To combat this effect it is necessary to arrange the circuit constants, particularly the resistance of the meter path, so that in all signal input conditions each of the transistors has approximately the same dissipation. With matched transistors VT5 and VT6 dissipate about 12 mW and they remain equal in this respect within 10 per cent over most of the signal range.

With no signal input the transistor bridge is unbalanced to a degree determined by the setting of the 'zero' control RV4 so as to bring the meter pointer to its left-hand zero position. When a signal is applied the bridge moves progressively towards balance causing the meter to read and the potential of VT5 collector to become more negative. At meter zero VT5 collector is more positive than the travellers of the 'law' controls RV2 and RV3, and the crystal diodes MR3 and MR4 do not conduct; but, as the collector goes negative with respect to each traveller, the corresponding diode passes current and negative feedback is applied to the base of VT5. The law controls are adjusted to make this action occur when the meter reading is about one-third and again at about two-thirds of full scale. The gain of the stage is therefore compressed in two successive steps resulting in an approximately logarithmic characteristic.

The crystal diode MR5 limits the meter reading to full scale if an overload signal should drive the transistor bridge through balance to a reverse unbalance condition.

The signal amplifier is similar to the output stages of the AM9/1 but with the output transistors connected in common collector in order to obtain a low output impedance which is further reduced by the feedback taken via the sensitivity control RV1. The resulting impedance at the emitters of about 100 ohms is modified to the required 312 ohms by the transformer and the resistance of the rectifiers MR1 and MR2.

The amplifier input circuit and the sensitivity are arranged to be suitable for measurement at the monitor output of the AM9/1 and other similar amplifiers.

Performance:

Sensitivity: Full scale deflection for 0·45 V peak input.

Scale accuracy: Within $\pm0\cdot5$ dB of the standard BBC meter calibration.

Drift: The measurement error arising from operating point drift of the final stage is within $\pm0\cdot5$ dB.

Selected, matched transistors for the final stage pair are needed to obtain the above performance.

4.3 *Microphone Pre-amplifier*

This amplifier forms part of a microphone channel assembly which also includes a channel fader with group and cue switches and various indicator lamps. The channel fader is a simple interstage gain control, R17, of a type similar to that used in the AM9/1, replacing the relatively expensive and bulky constant impedance fader necessarily used with the equivalent valve apparatus.

The first two stages (Fig. 5) form a feedback pair with both series and shunt a.c. feedback, and also d.c. feedback taken from a resistance in the emitter circuit of the second stage. The relative magnitudes of series (from R10 via C4 and the transformer secondary) and shunt a.c. feedback (from R9 via R1 and C2) are adjusted to give the required 600-ohm amplifier input impedance with the input transformer ratio required for minimum noise conditions. Together they have a magnitude of 24 dB. The d.c. conditions are well stabilized by the d.c. feedback from R8 via R11. The collector currents of the two stages are:

$$i_{c1}=0\cdot31+1\cdot15\, i_{co1}+0\cdot15\, i_{co2} \text{ mA}$$
$$i_{c2}=0\cdot77-1\cdot3\, i_{co1}+1\cdot4\, i_{co2} \text{ mA}$$

The output stage is an orthodox arrangement with current feedback and a collector load giving the required 600-ohm unbalanced output impedance. The current swing in the output transistor is relatively small and adequately low distortion is obtained without the complication of a push-pull circuit.

Performance:

Source impedance 300 ohms
Load impedance 600 ohms

Maximum insertion gain at 1 kc/s 46 dB

Frequency response:

60 c/s to 10 kc/s within 0 to $-0\cdot5$ dB
40 c/s to 15 kc/s within 0 to $-1\cdot0$ dB

Non-linearity:

Total harmonic distortion for normal peak output of -10 dBm:

at 1,000 c/s $<0\cdot1$ per cent
at 60 c/s $<0\cdot2$ per cent

For $+10$ dBm output: $\not>1\cdot0$ per cent

Noise factor: 5 to 7 dB—depending on first transistor —at a bandwidth of 400 c/s to 10 kc/s

Terminal impedances at 1 kc/s:

$|Z \text{ out}|=600$ ohms ±5 per cent
$|Z \text{ in }|=600$ ohms ±10 per cent

Power consumption: 12 mA at 24 V.

4.4 *Line-sending Amplifier*

This amplifier falls into the Group (*b*) referred to in Section 2, and the performance requirements are more stringent than those of the amplifiers already described. It is a bridging amplifier with a high-impedance input cir-

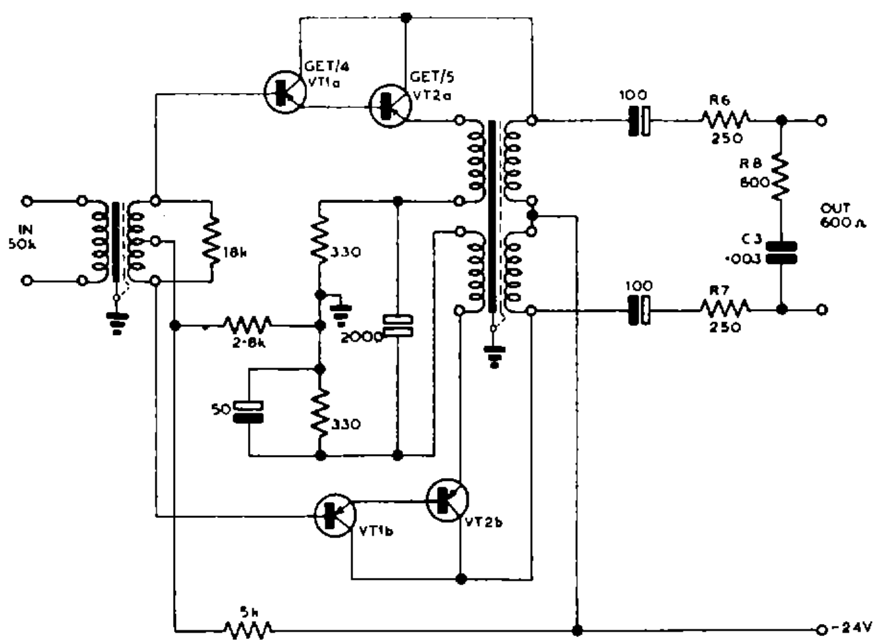

Fig. 6 — Line-sending Amplifier: Circuit

cuit, and the gain is fixed at 10 dB. The circuit is given in Fig. 6.

By the use of a push-pull pair of the 'super-alpha' composite arrangement of Fig. 1 sufficient gain is obtained from what is effectively a single stage to provide the amount of feedback necessary for the high performance required. The feedback is obtained from a winding on the output transformer and has a magnitude of 24 dB.

The impedance, looking into the base circuits of the two first transistors, is approximately 0·25 megohms in parallel with 200pF. and the input transformer is designed to give a flat frequency response when working into this impedance in parallel with its load resistance R1 of 24,000 ohms. The amplifier input impedance is 50,000 ohms.

In addition to a flat frequency response and low distortion, it is important, because of the requirements of line equalization, that the output impedance should be an accurate 600 ohms of very small angle and also that the frequency response should be independent of level to a high degree.

It is necessary for the output circuit to be balanced, but because in use the amplifier is followed by an accurately balanced line transformer, a balanced choke output with isolating condensers is adequate. This arrangement avoids the effects of the leakage inductance necessarily present in an output transformer. These effects are serious in this application for a transformer of reasonable construction because of the exacting requirements with regard to frequency response and output impedance. The core of the choke is of mu-metal with a small air gap in order to avoid any significant change of primary inductance with signal level. An inductance which varies in this way may cause a phase change in the feedback loop at low frequencies of sufficient magnitude to affect the frequency response with respect to signal level. It is important that this effect does not occur with this or with the line-receiving amplifiers.

The output impedance at the output choke is 100 ohms in series with approximately 1·2 mH. This impedance is built out to a constant resistance network of 600 ohms by R6 and R7, together with the reactive shunt arm R8 and C3.

Expressions defining the d.c. conditions of a 'super-alpha' pair are given in the appendix. For the constants of this amplifier they become:

$$i_{c1}=0\cdot44+1\cdot1\ i_{co1}-i_{co2}\quad \text{mA}$$
$$i_{c2}=26+5\cdot4\ i_{co1}+1\cdot1\ i_{co1}\ \text{mA}$$

Since a single potential divider is used to supply bias to both of the transistor bases the two pairs are d.c. coupled to a small degree, but this has an insignificant effect on the operating conditions.

Performance:

Source resistance 300 ohms
Load resistance 600 ohms

Insertion gain at 1 kc/s: $10\pm0\cdot2$ dB

Frequency response:

40 c/s to 10 kc/s within $\pm$ 0·15 dB
25 c/s to 20 kc/s within $\pm$ 0·25 dB
relative to 1 kc/s

Non-linearity:

At normal peak output of +18 dBm the total harmonic content is less than 0·2 per cent between 60 c/s and 5 kc/s and similarly at +21 dBm output is less than 0·5 per cent.

Output impedance:

The return loss against 600 ohms is better than 35 dB between 60 c/s and 15 kc/s.

Power consumption:

56 mA at 24 V.

4.5 *Line-receiving Amplifier*

This amplifier is intended to follow the line equalizer at BBC centres and to make good the line and equalizer losses. It must therefore have the good general performance of Group (*b*) amplifiers, in particular an accurate 600-ohm input impedance, and a fairly high gain adjustable over a wide range.

From the circuit diagram (Fig. 7) it will be seen that the general scheme of two feedback pairs is used here as before. VT1 and VT2 are connected as a d.c. and a.c. feedback pair in the manner of the microphone pre-amplifier, but with only the series connected a.c. feedback—derived from R6 and RV2 and fed to the base of VT1 by way of C3 and the transformer secondary circuit. Since shunt feedback is not used the impedance at the base of VT1 is high and the required input impedance is obtained from the resistive load RV1 on the input transformer. A more accurate input impedance can be obtained in this way than by the use of combined series and shunt feedback, at the expense of a slightly worse noise factor, which will be higher by not more than 3 dB. This latter effect is of no consequence in an amplifier of this type where the level at the input to the first transistor will not normally be less than −55 dB, and the signal/noise ratio may, therefore, be made quite adequately high even with the above limitation.

The effect of the transformer leakage inductance on the input impedance is reduced by the condensers C1 and C2, which with the leakage inductance form an approximation to a low-pass filter section having a cut-off frequency of about 50 kc/s. The resistance RV1 is arranged as an input gain control, variable in 10 dB steps, and the feedback over the two stages may be varied from a magnitude of 22 dB to 32 dB by means of RV2, which forms a fine gain control having half-dB steps over a 10-dB range.

The push-pull output stage is similar to that of the line-sending amplifier previously described, but a normal output transformer is required here; this does not lead to any difficulty as the output impedance requirements are not so stringent as before. Feedback, of a magnitude of 18 dB, is taken from a tertiary winding of the output transformer. The input and output stages are coupled by the phase-splitter VT3 which is arranged to feed equal signal currents to each half of the output stage from equal impedances.

The d.c. conditions of the output stage are very similar to those of the line-sending amplifier, except that the output transistors are run at a lower dissipation since less power output is required. The two input stages closely re-

semble the corresponding part of the microphone pre-amplifier described in Section 4.3.

Performance:

Source impedance	600 ohms
Load impedance	600 ohms

Insertion gain at 1 kc/s: Variable from 20 dB to 60 dB in 0·5-dB steps

Frequency response:
From 40 c/s to 10 kc/s within $\pm 0 \cdot 15$ dB
From 30 c/s to 15 kc/s within $\pm 0 \cdot 3$ dB
relative to 1 kc/s

Non-linearity:
The total harmonic distortion between 60 c/s and 5 kc/s does not exceed 0·3 per cent at normal peak output of +8 dBm, and 0·5 per cent at an output of +16 dBm.

Terminal impedances:
The return loss of the input impedance is not less than 35 dB at 1 kc/s and 25 dB between 60 c/s and 15 kc/s.
$|Z \text{ out}| = 600$ ohms ± 10 per cent 60 c/s to 10 kc/s

Noise:
Noise factor 7 to 9 dB for a bandwidth of 400 c/s to 10 kc/s

Power consumption:
44 mA at 24 V.

5. Mechanical Construction

Most of the amplifiers are made up of two printed-circuit assemblies containing the small components. All except the microphone pre-amplifier have a multi-pin plug at the rear for connection to their associated circuits and external power supply, where this is used, and are held in position by Dzus quick-release fasteners. The appearance and internal construction of the amplifiers are illustrated in Figs. 8 and 9.

5.1 *AM*9/1 *Amplifier and ME*12/1 *Programme Meter*

These are constructed in a book-shape chassis of aluminium alloy of main body dimensions $7\frac{5}{8}$ in. $\times$ $5\frac{7}{8}$ in. $\times$ $2\frac{1}{2}$ in., and front panel 7 in. $\times$ $2\frac{7}{8}$ in. They are normally used in groups and take power from a mains-driven 24-V power unit, but for other applications space is provided inside the amplifier for a mercury-cell battery capable of driving it for fifty hours.

5.2 *Microphone Pre-amplifier*

The dimensions are $3\frac{1}{4}$ in. $\times$ 3 in. $\times$ $1\frac{3}{8}$ in., with the input transformer protruding at the bottom. The construction was arranged to fit into the fader panel assembly, illustrated in Fig. 8. This panel is intended for mounting in groups in a studio control-desk or similar equipment which also contains one or more additional amplifiers of the same type as group amplifiers, together with an output amplifier similar to the line-receiving amplifier described above and the necessary power-supply unit. The low power dissipation and low microphony of transistors make them particularly adaptable to self-contained desks of this nature.

5.3 *Line-sending and Line-receiving Amplifiers*

These amplifiers are constructed on the same type of chassis as the present standard valve amplifiers.[2] They may therefore be assembled on the same type of mounting panel and so become integrated into existing systems.

6. Conclusions

During the course of design work leading to these amplifiers and others of a similar nature, some of the limitations and advantages offered in this field by the transistors at present available have become evident. They may be summarized as follows:

6.1 *Apparatus Size*

Much attention has been given generally to the small size of apparatus made possible by transistors. This reduction of size may be fully realized only where the valves otherwise used form a major part of the bulk of the apparatus or where the performance required is low enough to enable economies to be made in associated components. These conditions do not in general occur in high-grade a.f. amplifiers. In these, the more complex circuitry commonly required by transistors is likely to absorb the space released by the valves, the transformers remain unaffected, and the amplifier may be no smaller than an equivalent valve amplifier of compact type.

However, the smaller and more simple power supplies, together with the much smaller heat production compared with valves, may have a great effect on the overall size of equipment assemblies, such as studio desks, and offer greater flexibility in constructional methods.

6.2 *Power Efficiency*

The much better power efficiency, maximum signal output/total power consumption, of a transistor amplifier compared with an equivalent valve amplifier may be seen by comparing the AM9/1 with the OBA/9,[1] a similar amplifier of economical design. For the transistor amplifier the total power consumption is 480 mW and the efficiency 10·4 per cent; for the OBA/9 consumption is 6,660 mW and the efficiency 0·75 per cent. A similar efficiency ratio exists between the other transistor amplifiers and their valve counterparts.

6.3 *Noise*

Transistors are now available which enable noise in audio amplifiers to be within about 6 dB of the thermal noise of the source impedance. The audible amplifier-hiss given by such an amplifier in extreme gain settings in microphone circuits is about 3 dB higher than that given by the best of equivalent valve amplifiers. This extra hiss may be imperceptible in some applications, but where low-output microphones are used for high-fidelity programmes the difference is sufficient to make the valve amplifier preferable.

6.4 *Other Components, Costs, and Reliability*

Because of the need to stabilize d.c. conditions and feedback circuits most of the resistors used in the amplifiers

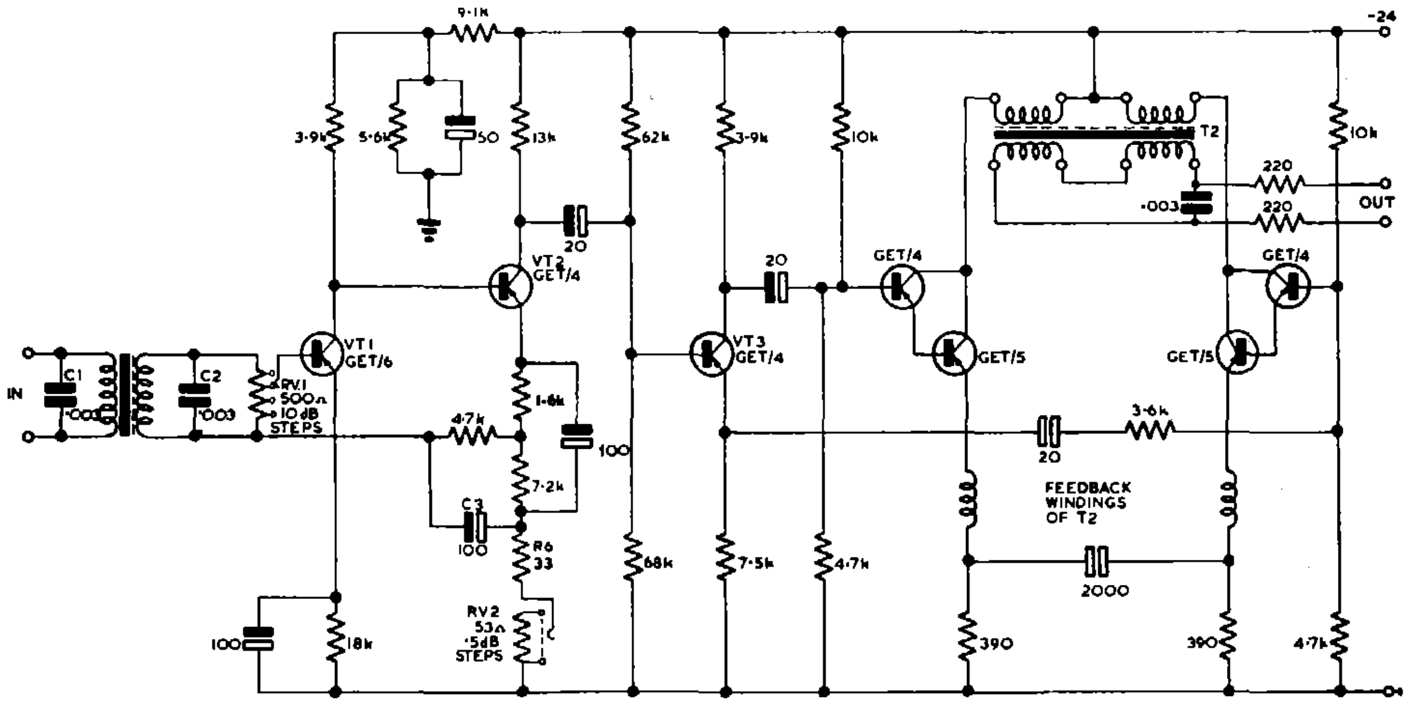

Fig. 7 — Line-receiving Amplifier: Circuit

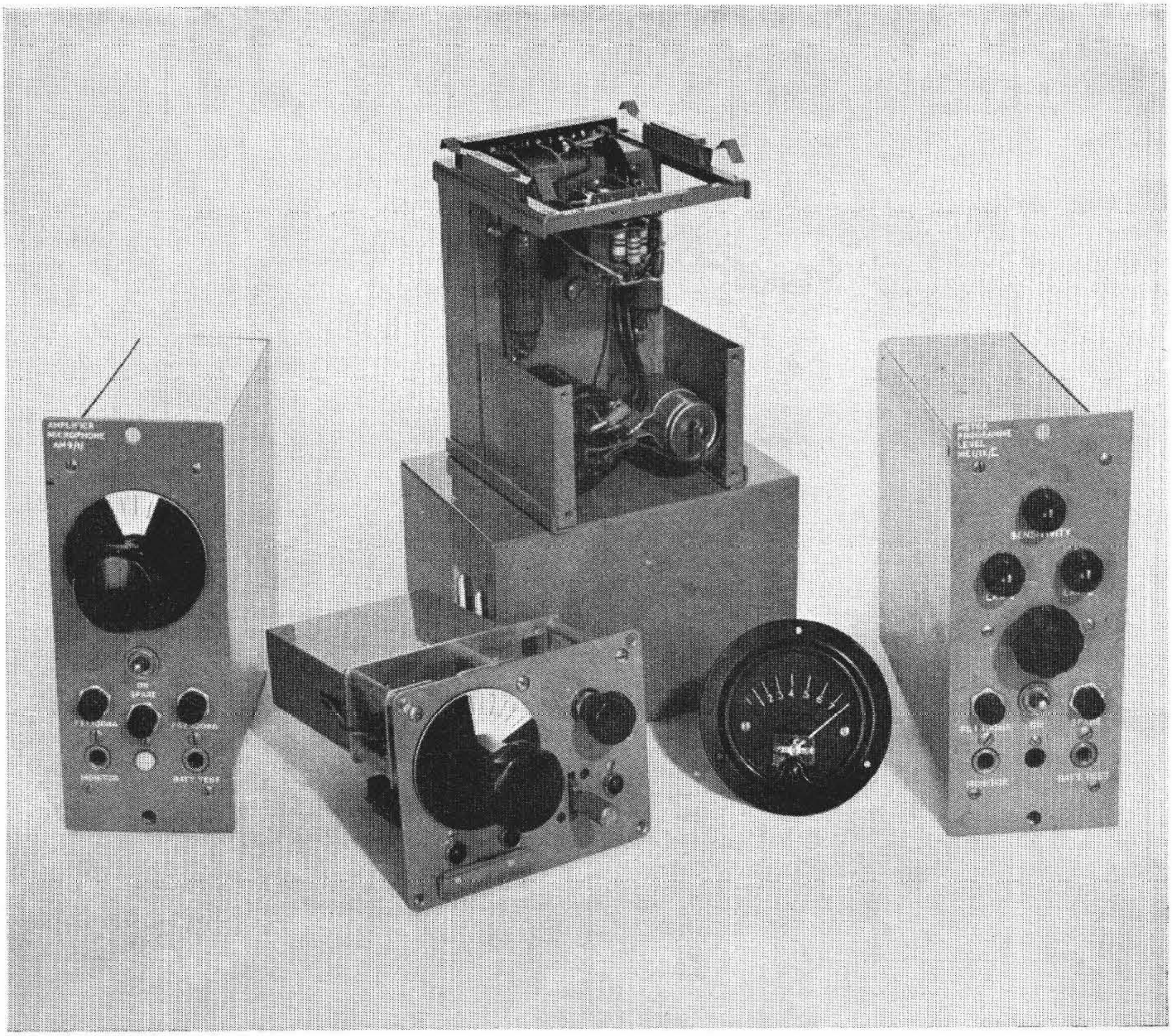

Fig. 8 — The group of amplifiers: the microphone pre-amplifier is in position at the rear of its associated fader panel. The line-receiving amplifier closely resembles the line-sending amplifier and only the latter is shown

described are high-stability types of ± 2 per cent accuracy. This is a common requirement of amplifiers of this nature. Also, a number of small high-capacity electrolytic condensers are likely to be needed, at least in multi-stage amplifiers. These factors, together with the fact that at present transistors are more expensive than valves and more of them are generally required, cause a transistor amplifier to be rather more expensive than an equivalent amplifier using valves.

The reliability of transistors, if a small proportion of early failures are rejected, is generally good, and may be expected to improve as production techniques advance. Life-times should be high and the available evidence points in this direction: like valves, transistors should be worked well within their maximum dissipation.

Less evidence is yet available of the reliability to be expected from the small electrolytic condensers. In most cases it should be possible to use them well within their temperature and voltage ratings, and considerable advances have been made in methods of sealing in recent years. It is possible, therefore, that a longer reliable life can be expected from them than from former types used in more trying conditions.

6.5 *Other Performance Characteristics*

The considerable variability in parameters, particularly in current-gain factor and cut-off frequency, shown by transistors at present prevents the achievement, at least in a design which can be consistently reproduced, of the highest grade performance of which valves are capable,

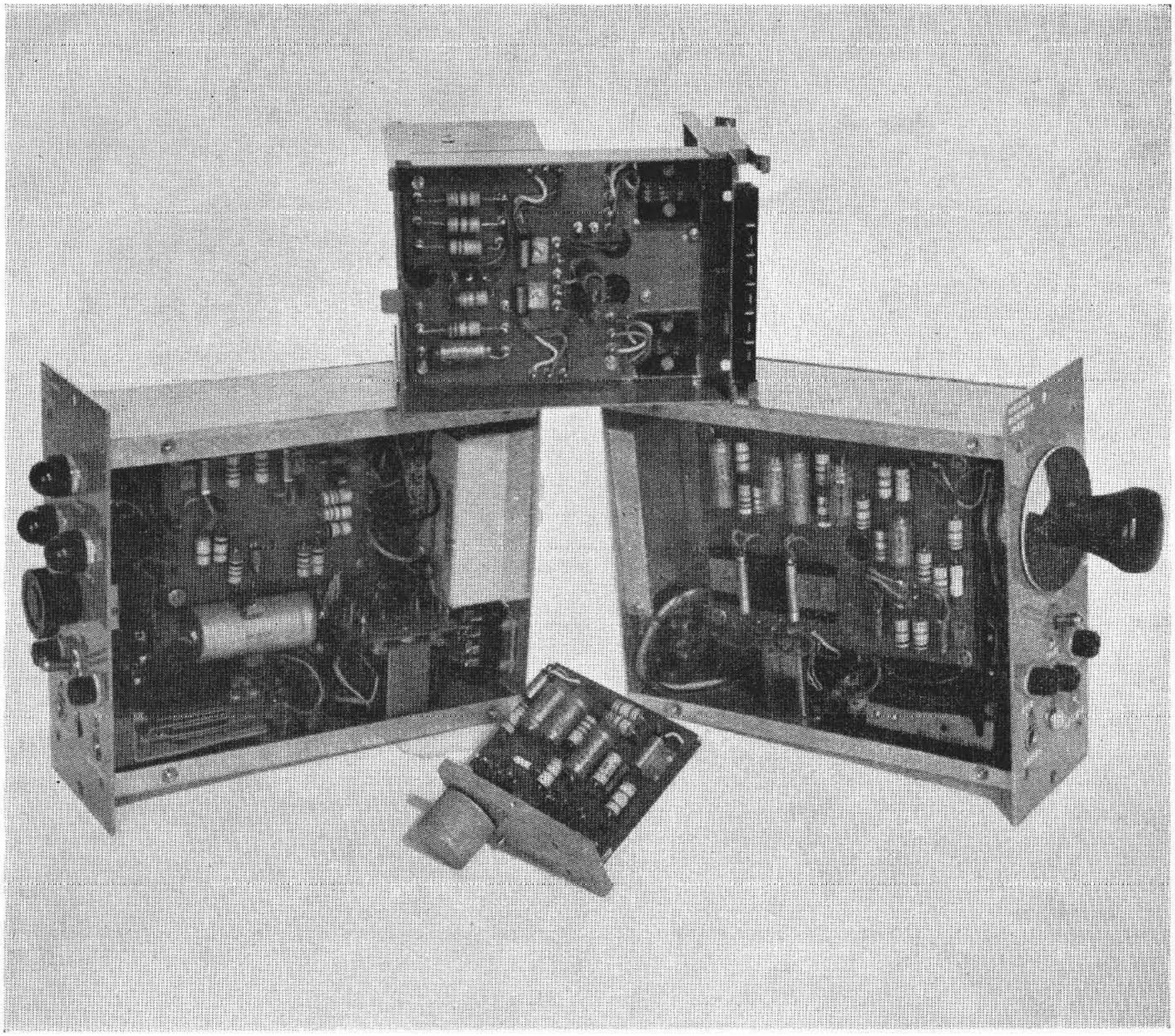

Fig. 9 — The interior construction of the amplifiers

unless a rigorous selection of transistors is made. This procedure is rarely practicable. Even for the slightly lower standards of the amplifiers described here a certain selection and rejection of extreme specimens is desirable. It is considered that, with regard to the best a.f. transistor types now available, greater consistency would be of more value at present than an improved average performance.

However, it is considered that a performance quite adequate for many applications in the field of high-quality sound reproduction is readily obtainable and that transistors offer many advantages in this sphere, even in the present state of development. As manufacturing techniques improve and types now under development become available the transistor should become an even more important rival to the valve.

7. Acknowledgments

The author would like to acknowledge the assistance received from his colleagues in the Designs Department during the work described, in particular from Mr G. A. McKenzie, who designed most of the transformers used in the amplifiers.

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APPENDIX

Operating Conditions of Transistor Pairs

The following assumptions are made:

Collector current is independent of collector voltage over the range of interest.

Base-emitter voltage is negligibly small.

α is constant over the range of interest and is high so that the following approximations may be made with very small error:

$$\alpha \approx 1$$

$$\frac{1}{1-\alpha} \approx \frac{\alpha}{1-\alpha} \approx \alpha'$$

and α' is the current gain in common emitter.

If i_c, i_e, i_b, and i_{co} refer to the total collector, emitter, base, and cut-off currents respectively, then:

$$i_e = i_b + i_c \quad \text{(1)}$$

$$i_c = \alpha'(i_b + i_{co}) \quad \text{(2)}$$

$$i_e = (1+\alpha')i_b + \alpha' i_{co} \quad \text{(3)}$$

(a) *The D.C. Feedback Pair*

Referring to Fig. 2 and with the current subscripts 1 and 2 denoting the first and second transistors:

$$V = R_1(i_{c1} + i_{b2}) + R_3\, i_{e2} + R_4(i_{e2} - i_{b1}) \quad \text{(4)}$$

$$0 = R_2\, i_{e1} + R_5\, i_{b1} + R_4(i_{b1} - i_{e2}) \quad \text{(5)}$$

Solving these equations for i_{c1} and i_{c2} results in:

$$i_{c1} = \frac{1}{D}\left[V + (A+B)i_{co1} + R_1\, i_{co2}\right] \quad \text{(6)}$$

$$i_{c2} = \frac{1}{C+R_4}\left[V - R_1\, i_{c1} - R_4\, i_{co1}\right] + \left[1 + \frac{R_1}{C+R_4}\right] i_{co2} \quad \text{(7)}$$

$$\text{where } C = R_3 + \frac{R_1}{\alpha_2'} \qquad D = R_1 + A + \frac{B}{\alpha_1'}$$

$$A = R_2\left[\frac{C}{R_4} + 1\right] \qquad B = C\left[1 + \frac{R_5}{R_4}\right] + R_5$$

Where two such pairs having the same resistance values are coupled together by means of a common emitter resistance R_2 as in the AM9/1 amplifier equation (4) still applies and (5) becomes:

$$0 = R_2(i_{e1} + i_{e1a}) + R_5\, i_{b1} + R_4(i_{b1} - i_{e2}) \quad \text{(8)}$$

where i_{e1a} denotes the emitter current of the first transistor of the additional pair.

Since the arrangement is symmetrical about R_2 the equations (4) and (8) may be solved to give expressions for the collector currents of all four transistors. In the more general case where the α' values of all four transistors may differ the resulting expressions are very cumbersome, but where the α' values of corresponding transistors are equal so that the arrangement is completely symmetrical they may be simplified to:

$$i_{c1} = \frac{V}{D+A} + \frac{1}{D^2 - A^2}\Big[(DA + DB - A^2)i_{co1} + R_1 D\, i_{co2} - A(A+B-D)\, i_{co1a} - AR_1\, i_{co2a}\Big] \quad \text{(9)}$$

$$i_{c2} = \frac{1}{C+R_4}(V - R_1\, i_{c1} - R_4\, i_{co1}) + \left[1 + \frac{R_1}{C+R_4}\right] i_{co2} \quad \text{(10)}$$

where the suffixes 1 and 2 refer to the first and second transistors of either pair and 1a and 2a refer to the corresponding transistors of the similar pair to which they are coupled. The other symbols have the same values as before.

(b) *The 'Super-alpha' Pair (Fig. 1)*

With the same assumptions as before the network equations are:

$$V = R_5(i_r + i_{b1}) + R_2\, i_{e2} \quad \text{(11)}$$

$$0 = R_2\, i_{e2} - R_6\, i_r \quad \text{(12)}$$

Solving these for i_{c1} and i_{c2} gives:

$$i_{c1} = \frac{1}{P + \frac{R_5}{\alpha_1'}}\left[V + (P + R_5)\, i_{co1} - P\, i_{co2}\right]$$

$$i_{c2} = \frac{1}{Q}(V + R_5\, i_{co1}) + \left[\frac{R_5}{Q\alpha_1'} + 1\right] i_{co2}$$

$$\text{where } P = R_2\alpha_2'\left[\frac{R_5}{R_6} + 1\right] \text{ and } Q = R_2\left[\frac{R_5}{R_6} + 1\right] + \frac{R_5}{\alpha_1'\alpha_2'}$$

The last term in Q is generally negligibly small, $P = \alpha_1' Q$, and the operating conditions of the second transistor are independent of α_2'.

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